

Please amend the paragraph on page 1, line 14 through page 2, line 4 as follows (this paragraph presented in the Appendix with amendments entered):

a<sup>2</sup>

When attempting to range on an emitting radar, the radar is seldom emitting a steady beacon for a significantly long observation interval. In fact, often such emitting radars are emitting for only a short time duration to avoid detection or other reasons. The short time duration is in the order of a hundred milliseconds or so. As will be discussed herein, the frequency difference observed by two spaced apart antennas at desired ranges [are] is often in the order of only a few Hertz in the desired ranges (about 20 kilometers). To achieve desired range accuracies (about 20% of range), it is necessary to not only measure the frequency difference, but also to measure the frequency difference very accurately. An improved approach is needed to achieve accurate enough measurements of phase rate in just a hundred milliseconds or less.

Please amend the paragraph on page 4, line 16 through page 5, line 2 as follows (this paragraph presented in the Appendix with amendments entered):

a<sup>3</sup>

In FIG. 2, an RF angle rate interferometer is constituted by a receiver that includes RF bridge 100 coupled through frequency converter 300 to processor 200. Frequency converter 300 up converts ([heterodynes] heterodynes up) processor reference signal 202 by a predetermined intermediate frequency step to form intermediate reference signal 302. In the example discussed herein the frequency step is 792 MHz so that intermediate reference signal 302 has a frequency of 800 MHz. In FIG. 2, RF bridge 100 produces an information signal that has the frequency difference between the frequencies of signals received at antennas 102 and 104 frequency modulated onto intermediate reference signal 302. Frequency converter 300 also down converts ([heterodynes] heterodynes down) the information signal (intermediate input signal 304) produced by RF bridge 100 by the predetermined intermediate frequency step (in the example discussed herein, 792 MHz) to form a

a<sup>3</sup>  
cont

down converted information signal that is used as processor input signal 204. In the example discussed herein, the down converted information signal that is used as processor input signal 204 is a signal having a frequency of 8 MHz plus any measured frequency difference between signals received at antennas 102 and 104. Processor 200 then computes this frequency difference and the corresponding angle rate of rotation.

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Please amend the paragraph on page 5, lines 22-25 as follows (this paragraph presented in the Appendix with amendments entered):

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a<sup>4</sup>

[Preferrably] Preferably, mixers 140 and 160 are single sideband (SSB) mixers that both produce the upper sideband mixer results. To the extent present in the outputs of mixers 140 and 160, any lower sideband signals, pump signals or other signals outside of the desired upper sideband are removed by filters 150 and 170, respectively.

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Please amend the paragraph on page 6, lines 5-9 as follows (this paragraph presented in the Appendix with amendments entered):

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a<sup>5</sup>

The signals from filters 150 and 170 are then combined in mixer 180. Mixer 180 is [preferrably] preferably a single sideband (SSB) mixer that produces the lower sideband mixer result. To the extent present in the output of mixer 180, any upper sideband signal or any other signal outside of the desired lower sideband is removed by filter 190 to provide the information signal from RF bridge 100.

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Please amend the paragraph on page 6, lines 21-28 as follows (this paragraph presented in the Appendix with amendments entered):

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a<sup>6</sup>

Signals from frequency source 110 and processor reference signal 202 are preferably spectrally pure. Typically, the frequency of processor reference signal 202 is based on a direct digital synthesizer, a frequency multiplied replica of a crystal

a<sup>6</sup>  
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oscillator, or a phase locked loop synthesizer. Filter 130 can be any narrow band filter (high Q filter) that operates in the frequency range (in this example, 80 MHz). For example, in a filter designed to pass a bandwidth of only 100 kHz, a Q of 800 would be required. Such filters include surface [accustic] acoustic wave devices (SAW devices) and some ceramic resonators, but there are many alternatives.

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Please amend the paragraph on page 7, lines 1-6 as follows (this paragraph presented in the Appendix with amendments entered):

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a<sup>7</sup>

Similarly, filters 150 and 170 are constructed from any narrow band filter (high Q filter) that operates in the desired frequency range (in this example, 134 and 126 MHz). Filters 150 and 170 are preferably designed to pass a bandwidth of 4 MHz (e.g., the +/- 2 MHz frequency range of the emitter signal), and such filters may be implemented with surface [accustic] acoustic wave devices (SAW devices) and some ceramic resonators, but there are many alternatives.

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Please amend the paragraph on page 8, lines 15-23 as follows (this paragraph presented in the Appendix with amendments entered):

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a<sup>8</sup>

Removal of the pump signal from the output of filter 130 is desired in order to prevent the pump signal from leaking through mixer 120 and through filter 130 to mix with a strong, off frequency, signal picked up by antenna 104. The pump signal is removed by either good isolation in single sideband mixer 120, or good band rejection by filter 130, or both. The filter "skirt" on the lower frequency end of the pass band may be required to drop 30 dB in just 325 MHz on the 4,620 MHz end of the bandpass. This is a 30 dB drop in just 7% of the bandwidth. Such a single filter will require at least 14 "polls" to achieve. Present filter technologies in the 4,600 to 5,100 MHz region use various types of stripline filters, microstrip filters, waveguide filters, coax filters, and the like.

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Please amend the paragraph on page 9, line 25 through page 10, line 6 as follows (this paragraph presented in the Appendix with amendments entered):

a<sup>9</sup>

In FIG. 4, processor 200 includes digital frequency source 206 and digital to analog converter 208 (DAC 208) to produce processor reference signal 202. Collectively, frequency source 206 and DAC 208 constitute a direct digital synthesizer. In the examples discussed above, processor reference signal 202 is a spectrally pure 8 MHz signal. Processor 200 further includes analog to digital converter 210 (ADC 210) to [received] receive the down converted information signal (FIG. 2) or the information signal (FIG. 1) as processor input signal 204. In the examples discussed above, signal 204 is an 8 MHz signal onto which has been modulated a frequency difference signal, the frequency difference signal being the difference in frequency between the emitter signal received at antenna 102 and the emitter signal received at antenna 104. This frequency difference is typically of the order of 1 Hertz under the circumstances describe in the examples herein.

Please amend the paragraph on page 10, lines 17-26 as follows (this paragraph presented in the Appendix with amendments entered):

a<sup>10</sup>

In FIG. 4, digital frequency source 230 provides a "local oscillator" signal having a frequency that is the sum of the frequency of processor reference signal 204 and an offset frequency. In the example discussed herein, the offset frequency is 64 Hz. The "local oscillator" signal is in digital form. Preferably, the "local oscillator" signal is a sinusoidal wave represented by complex digital numbers at a predetermined sample rate that matches the sample rate at the output of filter 220. In the present example, ADC 210 samples its input signal at 16 MSPS (million samples per second) and filter 220 filters the signal using a 16 MHz clock frequency to process the 16 MSPS from ADC 210 through filter 220. The "local oscillator" signal is a digitally sampled 8 MHz sinusoidal wave that is sampled at 16 MSPS. Preferably, all digital representations of the various signals are complex numbers.

Please amend the paragraph on page 22, lines 8-18 as follows (this paragraph presented in the Appendix with amendments entered):

a<sup>11</sup>  
The DFT integration interval is still preferably one-half  $\pi$  divided by the displacement frequency defined by  $\delta\omega$  but in Hertz (the [difference] difference between the center frequency of the DFT and the offset frequency defined by  $\omega_0$ ) as discussed above with respect [ot] to equations (4), (5) and (6). In the example discussed herein,  $\delta\omega$  is  $\pi$  divided by 64, or 49 milliseconds. If the dwell time of the EMITTER is only 40 milliseconds, then the remaining 9 milliseconds of the DFT integration interval is filled with a balanced number of leading and trailing zero values for calculation purposes (so that the shape of the two discrete Fourier transforms combine to form a linear frequency discriminator). In the case of a 40 millisecond dwell time, the required signal to noise ratio is determined by Equation (16) as if T were 40 milliseconds even though the DFT filter function uses a 49 millisecond integration interval. The required SNR is 312.5 (24.95 dB).

Please amend the paragraph on page 23, lines 9-18 as follows (this paragraph presented in the Appendix with amendments entered):

a<sup>12</sup>  
In an exemplary EMITTER modeled after the description in the The Radar Handbook, second edition, published by McGraw Hill, Merrill Skolnik as editor, 1990, page 7.72, the EMITTER has an antenna gain  $G_T$  of about 6,250 (i.e., 37.95 dB), a peak power of 3 megawatts and an average power of 5 kilowatts. The pulse [repretition] repetition interval PRI is 600 times the pulse duration PD. In Equations (17) and (19),  $P_T$  may be regarded as the average power (i.e., 5 kilowatts or 66.99 dBm referenced to one milliwatt). If  $\lambda$  is 0.1 meters (i.e., based on 3 GHz) and  $G_R$  is taken to be 2 dB, then  $S_R$  is [-75.07] -15 dBm (i.e., about [0.03 microwatt] 32 milliwatts). With a required SNR of 24.95 dB (based on a 40 millisecond dwell time), the noise level should not be allowed to rise above [-100] -40 dBm. In the exemplary embodiment described herein, the noise level is kept below -100 dBm.

Please amend the paragraph on page 24, lines 10-17 as follows (this paragraph presented in the Appendix with amendments entered):

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a<sup>13</sup> One source of this difficulty is the short dwell times that are available when intercepting a scanning antenna EMITTER. In the example herein, the dwell time is 40 milliseconds. Typically, this dwell time is parsed into three to five look times, for example, four look times averaging 10 milliseconds each. During each look time, the EMITTER transmits pulses repeated at a different pulse [repetition] repetition interval (i.e., inverse of pulse [repetition] repetition frequency PRF) so as to resolve ambiguous [doppler] Doppler indicated velocities in low and medium PRF modes and so as to resolve [ambiguous] ambiguous range reflections in medium and high PRF modes.

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Please amend the paragraph on page 24, lines 18-28 as follows (this paragraph presented in the Appendix with amendments entered):

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a<sup>14</sup> When switching from one look to another, the EMITTER frequency may also be switched. The angle rate interferometer described herein is not adversely effected by this form of frequency switching because it is only the frequency difference between the frequencies received at antennas 102 and 104 that is measured. Because of the pulse structure of the EMITTER's waveform, it is not possible to receive one frequency at antenna 102 and another frequency at antenna 104. On the occasional time when the measurement (ADC sampling) exactly corresponds to a time when one antenna receives signal power and the other antenna receives no signal power, the described angle rate interferometer treats the measurement as noise. If one were concerned with this [occurrence] occurrence, a signal power level threshold circuit may be installed at the outputs of mixers 140 and 160 to detect this condition and block the ADC sampling.

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